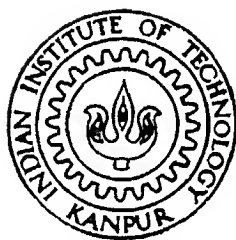


Digital LOS Studio – Transmitter Link for Doordarshan

by
Archana Dighe



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1997

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DEPARTMENT OF ELECTRICAL ENGINEERING

DIG, INDIAN INSTITUTE OF TECHNOLOGY KANPUR

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Digital L O S Studio - Transmitter Link for Doordarshan

A Thesis Submitted

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for the Degree of

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by

Archana Dighe

to the

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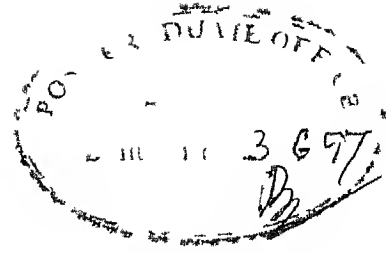
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This is to certify that the work contained in the thesis entitled **Digital L O S Studio Transmitter Link for Doordarshan** by **Archana Dighe** (Roll No 9612402)has been carried out under our supervision and that this work has not been submitted elsewhere for a degree

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Archana Dighe

Abstract

The aim of this work was to study different aspects of point to point digital T V transmission over the line of sight link

Presently in the Doordarshan network, all the line of sight links (Studio Transmitter O B spot Studio, uplinks and downlinks) are analog based on frequency modulation technique

In this thesis a modification of the existing analog L O S Studio Transmitter link at Jaipur, by a Digital link is proposed. As this is a changeover from analog to digital system it would be using some of the subsystems of the existing analog version. The parameters like frequency of operation, bandwidth allocation, receiver noise figure are also retained. It is presumed that encoded data is available at the transmitter input. Other parameters needed for the system design are taken from CCIR standards.

In the first part, system design is done which includes link budget calculations. In particular, the design is suitable for the climate and terrain at Jaipur and similar places. But necessary data for other types of terrains and atmospheric conditions is also given.

In the second part, the frequency of operation is chosen as 20 GHz, which is a less occupied band so that there is no bandwidth constraint.

In the end, a brief description of the subsystems is given.

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Chapter 1

Introduction

The migration from a television service dependent primarily on the application of analog technologies to one that is based on digital technologies has been evolving over the past thirty years. This television service migration is part of a natural outgrowth of the convergence of the television, telecommunications and computer arts and sciences through the shared use of digital technology.

The input and output signals of television at the camera and at the receiver respectively, are inherently analog. Thus, the question 'why digital?' is a natural one.

1.1 Drawbacks of Analog Video Systems

1. It is always difficult to maintain throughout the broadcast chain a true analog of the original signal. For example, the analog video consists of frequency components varying from DC to about 5 MHz. The wide frequency band has to be processed through a number of circuits before being transmitted. Since it is very difficult to design circuits which will have equal response to all of these frequencies, different circuits in the chain will contribute different distortions to the video waveform. Due to this, in analog systems the output signal is not always analogous to the input signals. A parameter can only be a true analog of the original if the conversion process is linear, otherwise harmonic distortion will result.
2. Distortion in phase and amplitude occur at all points in the system causing dete

rioration of the technical quality of the picture. For example, group delay caused in the circuit due to the difference in the velocities of propagation of different frequency components in the medium giving a phase distortion and the amplitude distortion due to the attenuation in certain stages in the circuit. Total degradation is the sum of all individual degradations of each circuit. This sets a limit on the number of stages.

- 3 Excessive attention is required in designing circuits so that the above distortions are limited to manageable proportions thus increasing the cost of the system.
- 4 In spite of the care taken by circuit designers and manufacturers, it remains a major maintenance problem to deal with the distortions that result from performance drift and lack of stability and keep equipment at all points in the programme chain operating to the standard required.
- 5 The effort necessary to maintain an acceptable standard is thus laborious and thus analog systems are costlier to operate and maintain to a given standard.
- 6 The analog signal cannot be stored beyond a certain small delay. As a few microseconds of a delay will involve lengthy delay lines. When it is required to delay the video by a few fields/frames as is the case of TBCs¹/frame synchronizers, one cannot estimate the size of the delay lines required for such an application.

1.2 Advantages of Digital Video Systems

The main advantages of the digital video are summarised as follows:

- 1 Signals can be processed and transmitted with a precisely calculated degree of degradation without any further random distortion. It means that the quality of a digital video link is independent of the characteristics of the channel in a properly engineered system. The frequency response, linearity and noise depends upon the conversion process.

¹Time Base Correctors

- 2 Since digital recording means the recording of bits, they can be copied through a large number of generations without degradation. This means that the life of the recording depends only on the quality of the storage medium since there is no degradation due to generation loss and the bits can be copied without any loss of information.
- 3 The use of error correction techniques eliminates the effects of dropouts and transmission errors.
- 4 It is possible to design and construct extremely precise and stable digital filters and equalizers with inherent phase linearity. Such devices need no adjustments, are much cheaper than analog ones and make complex processing very simple.
- 5 While signal degradations in the analog signal are cumulative and the characteristics of the degradations make them difficult to distinguish from the video signal, the ability to regenerate a digital pulse train exactly renders the digital signals theoretically immune to impairments from the external sources.
- 6 Digital bit streams can be interleaved within a single channel. This interleaving process allows for the emission, transmission, storage or processing of ancillary signals along with the video and associated audio.

1.3 Statement of the Problem

The present Studio Transmitter Links (STLs) in the Doordarshan network are analog type, using *Frequency Modulation* technique, with operating frequency 6 GHz. In this thesis, a *digital version* of the STL has been proposed and the necessary system design is done for the following two cases:

- (1) 6 GHz link (Modification of the existing analog link)
- (2) 20 GHz link (New link)

1 4 Specifications of the System

Following are the specifications of the proposed system

Frequency of operation CASE I 6 GHz

CASE II 20 GHz

R F Bandwidth 30 MHz for 6 GHz link

No Bandwidth constraint for 20 GHz link

Probability of bit error 10^{-7}

Availability of channel 99.99%

Bit rate 34 Mbps

The link is one way and single hop

1 5 Organisation of Thesis

Chapter 2 explains the factors that are to be considered while designing the system. This includes specifications of the existing system to be used in this system design, losses to be incorporated and the necessary assumptions.

Chapter 3 deals with the proposed digital modulation schemes for the two systems, i.e., 6 GHz and 20 GHz links. Brief description of the schemes with their error probabilities is also given in this chapter.

The link budget calculations for the 6 GHz and 20 GHz links, using the expressions and specifications from Chapters 2 and 3, are given in Chapter 4.

The brief description of the subsystems and specifications of the transmitter and receiver are included in Chapter 5.

Chapter 6 presents the results in a tabular form and also concludes the thesis giving scope for further work in related areas.

Chapter 2

Factors Influencing System Design

In the design of the proposed system, some of the specifications are used from the existing analog system. Other required specifications are mentioned in Sec. 1.4. Before going for the system design, the losses in the system must be known. These losses are discussed in Sec. 2.2 for both cases, i.e., 6 GHz and 20 GHz links.

2.1 Specifications

The following specifications of the existing analog system are retained for this system design.

The L O S distance between transmitting and receiving antenna is 10 km. Height of the transmitting antenna is 462 m and that of the receiving antenna is 596 m. This gives the beam depression angle of the receiving antenna roughly equal to 0.8° and the antenna temperature, corresponding to this angle is 140 K at 6 GHz operating frequency and 300 K at 20 GHz, [1] (pp. 2.9).

The bandwidth allocated for the channel is 30 GHz.

Both the transmit and the receive parabolic dish antennas have diameters of 2.2 m. Thus, the gain of each antenna is [8]

$$G = \frac{\eta 4\pi A}{\lambda^2} \quad \text{dB}$$

where η is the efficiency of the parabolic antenna and A is the aperture area

$$G_t = G_r = 40.24 \text{ dB} \quad (2.1)$$

The total gain $\Sigma G = G_t + G_r$ dB

$$= 80.48 \text{ dB} \quad (2.2)$$

If the same antenna is used for 20 GHz link the gain of each antenna would be 50.7 dB and the total gain is

$$= 101.4 \text{ dB} \quad (2.3)$$

2.2 Losses in the System

In this section, the losses significant for our system are considered

- The starting point for radio system path calculations is the transmission of a signal from a lossless isotropic transmitting antenna to a lossless isotropic receiving antenna in free space. Its effective aperture area is $\frac{\lambda^2}{4\pi}$ where λ is the wavelength of the signal. The ratio of the transmitted power to the received power for antennas separated by many wavelengths is the *free space loss* L_{fs} , which is given by

$$L_{fs} = 20 \log\left(\frac{4\pi d}{\lambda}\right) \text{ dB} \quad (2.4)$$

where the distance d between antennas is measured in the same units as λ

$$L_{fs} = \begin{cases} 128 \text{ dB} & \text{at 6 GHz} \\ 138.5 \text{ dB} & \text{at 20 GHz} \end{cases} \quad (2.5)$$

- Waveguide loss. There are three basic types of waveguides: Rigid rectangular, Rigid circular, and Semiflexible elliptical. Rigid circular waveguides have the lowest loss. But they are used mostly in straight runs and require complex networks for circular to-rectangular transitions. Rectangular and semiflexible elliptical waveguides have

similar characteristics They are widely used because they are easy to install in long continuous runs

The waveguide run of the existing system is to be used in the new system The *wave guide losses*¹, G_1 are 4 dB each at the transmit and receive the end For 20 GHz system WR 75 waveguide can be used which would introduce a loss of 9 dB each at the transsitter and at the receiver

- Equation 2.4 gives the free space loss, i.e., in the absence of atmospheric² or terrain effects It is essential to provide a *fade margin* in the design of the system so that it will continue to operate satisfactorily in the presence of multipath and dispersive fading³

If availability of channel, $A = 99.99\%$, then

the outage time, $T = 3153.6$ sec/year

T is given by

$$T = 79abfd^3(10^{-\frac{F}{10}} + 10^{-\frac{F_d}{10}}) \quad (2.6)$$

The above formula is from [1] (pp. 4.9) and is based on an empirical model, where, F = flat fade margin, F_d = dispersive fade margin and

$$a = \begin{cases} 1 & \text{for average terrain} \\ 0.5 & \text{for dry mountain terrain} \\ 4 & \text{over water terrain} \end{cases} \quad (2.7)$$

$$b = \begin{cases} 0.25 & \text{for average terrain} \\ 0.125 & \text{for dry mountain terrain} \\ 0.5 & \text{over water terrain} \end{cases} \quad (2.8)$$

¹This includes the losses in the entire RF chain in the transmitter and the receiver

²below 10 GHz atmospheric effect e.g. attenuation due to rain can be neglected

³This effect is due to dispersion hence the digital radio systems are said to have

Dispersive Fade Margin

F can be calculated using the following relation from [2] (pp 61)

$$T = \begin{cases} 8 \times 10^{-7} f d^{2.5} \times 10^{\frac{-F}{10}} & \text{for average terrain} \\ 16 \times 10^{-7} f d^2 \times 10^{\frac{-F}{10}} & \text{for dry mountain terrain} \\ 2 \times 10^{-7} f d^3 \times 10^{\frac{-F}{10}} & \text{over water terrain} \end{cases} \quad (2.9)$$

Equations 2.6, 2.7, 2.8 and 2.9 give the *flat* (F), *dispersive* (F_d) and *total fade margin* (FM) as given below

Terrain	F dB	F _d dB	FM dB
Average	11.8	13.5	26
Dry Mountain	9.82	1.06	11
Water	12.96	14.87	28

Table 2.1 Fade Margin

The Total Loss $\Sigma L = L_f + 2G_f + FM$

$$= \begin{cases} 162 \text{ dB} & \text{for average terrain} \\ 117 \text{ dB} & \text{for dry mountain terrain} \\ 164 \text{ dB} & \text{over water terrain} \end{cases} \quad (2.10)$$

- At frequencies above 10 GHz rainfall can cause enough attenuation on a microwave path to require consideration in the system design. The typical values of attenuation for heavy (50mm/hr) and moderate (20mm/hr) rainfalls are 5 dB/km and 2 dB/km respectively at 20 GHz operating frequency [1] (pp 4-12)
 - The 20 GHz system is interference limited where the following could be the sources of interference (1) a cross polarised channel (2) an adjacent copolarised channel and (3) reception of a frequency on a hop facing the opposite direction of the hop being interfered with.
- As the link under consideration is a single hop one way link and no other adjacent

copolarised or cross polarised channel exists the above sources of interference need not be considered Hence the total loss⁴ is given by

$$\sum L = \begin{cases} 176.5 \text{ dB} & \text{For moderate rainfall} \\ 206.5 \text{ dB} & \text{For heavy rainfall} \end{cases} \quad (2.11)$$

- Degradation in $\frac{E_b}{N}$ For a real system design, practical value of *Bit Energy to Noise Density Ratio*⁵ $\frac{E_b}{N}$ will be required, which differs from the ideal value by an amount called *degradation* Table 2.2 gives the details of degradation budget [2]

Cause	Degradation (dB)
Phase & amplitude errors of modulator	0.1
ISI caused by filters	1.0
Carrier recovery phase noise	0.1
Differential encoding/decoding	0.3
Jitter(imperfect sampling instants)	0.1
Excess noise bandwidth of demodulator	0.5
Other hardware impairments(temp. variations etc)	0.4
AM/PM conversion of the quasilinear output stage	1.5
Group delay	0.3
Feeder echo distortion	0.2
Total degradation(with/without differential coding)	4.5/4.2

Table 2.2 Degradation budget for the link

The 6 GHz link is mainly noise limited The factors to be considered while designing a system are free space path loss, losses in the RF components of transmitter and receiver, the flat fade and dispersive fade margin, degradation in $\frac{E_b}{N}$, the antenna gains and noise temperatures of antenna and receiving system

⁴ L_{fs} + waveguide loss + attenuation due to rain

⁵For explanation refer to Chapter 3

As mentioned above the 20 GHz link is interference limited. In the absence of any interfering channels, the factors that influence system design are free space path loss, losses in the RF chain at transmitter and receiver, attenuation due to rainfall degradation, in $\frac{E_b}{N}$, the antenna gain and noise temperatures of the antenna and the receiving system.

Chapter 3

Proposed Digital Modulation Schemes

A digital modulator and a demodulator are integral parts of digital radio transmitter and receiver. Digital modulation techniques which may be used to modulate the carrier may modify the amplitude, the phase or the frequency of a carrier in discrete steps. Performance analysis of the techniques depends primarily on the type of demodulation or detection used in the receiver. There are two major classes of detection: *coherent* or synchronous detection which is used in *phase shift keying* (PSK) and assumes that in the demodulator local carrier waveforms have a fixed phase relationship with that of the transmitted carrier. This adds to the complexity of the receiver circuitry. The other class is that of *non-coherent* or envelope detection which is normally used in conjunction with *amplitude shift keying*(ASK) and *frequency shift keying*(FSK). In the following two sections different digital modulation schemes and the probability of error for each are discussed. In the subsequent two sections schemes suitable for the systems under consideration are proposed.

3.1 Digital Modulation Schemes

In this section various digital modulation schemes are explained in brief. Out of these, schemes suitable for the two systems, i.e. 6 GHz and 20 GHz links are proposed in

3 1 1 Binary Phase Shift Keying

In *binary PSK* the two signalling waveforms may be represented by

$$s_1(t) = \sqrt{\frac{2E_b}{T_b}} \cos \omega_c t \quad 0 \leq t \leq T_b \quad (3.1)$$

$$s_0(t) = \sqrt{\frac{2E_b}{T_b}} \cos(\omega_c t + \pi) \quad 0 \leq t \leq T_b \quad (3.2)$$

where, $s_1(t)$ represents the binary 1, $s_0(t)$ represents the binary 0, T_b is the bit interval and E_b is the bit energy s_1 and s_0 are 0 elsewhere The signal transmission encoding is performed by a *nonreturn to zero*(NRZ) level encoder The resulting binary wave (in polar form) and a sinusoidal carrier are applied to a product modulator as shown in Fig 3 1(a) The carrier and the timing pulses used to generate the binary wave are usually extracted from a common master clock The desired PSK wave is obtained at the modulator output

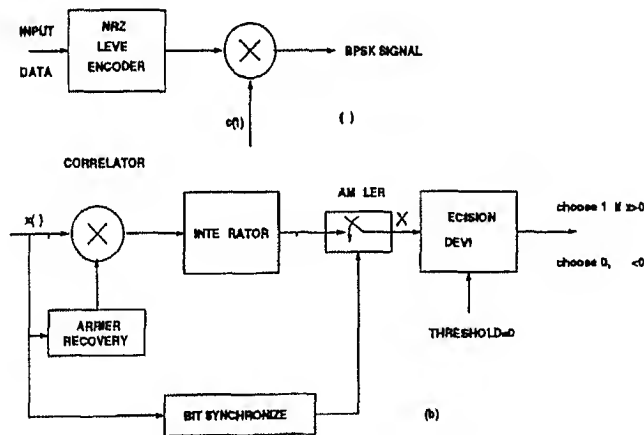


Figure 3 1 (a)Modulator, and (b) Demodulator for BPSK

To detect the original binary sequence of 1s and 0s, the received noisy BPSK signal $x(t)$ is fed to a correlator which is also supplied with a locally generated coherent carrier and recovered clock, as shown in Fig 3 1(b) The correlator output x_1 is compared with a threshold of zero volts If $x_1 \geq 0$, the receiver decides in favour of symbol 1, otherwise in favour of symbol 0

3 1 2 Quadriphase - Shift Keying

The provision of reliable performance, exemplified by a very low probability of error, is one important goal in the design of a digital communication system. Another important goal is the efficient utilization of channel bandwidth. In *quadriphase shift keying (QPSK)* as with BPSK, information carried by the transmitted signal is contained in its phase. In particular, the phase of the carrier can take one of the four equally spaced values, such as $\pm 45^\circ$, $\pm 135^\circ$. Fig 3 2 shows the signal constellation for QPSK.

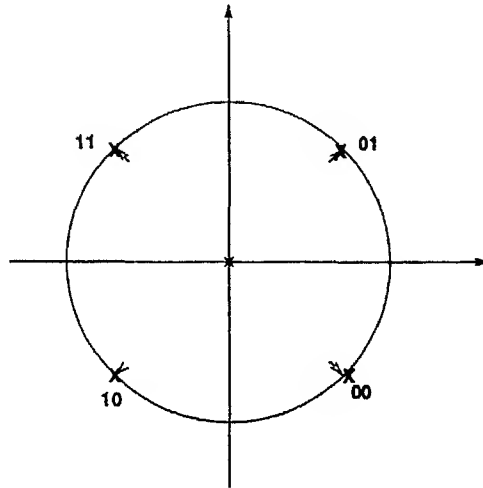


Figure 3 2 Signal Constellation for QPSK

Hence, the transmitted signal is given as

$$s_i(t) = \begin{cases} \sqrt{\frac{2E}{T}} \cos \left[2\pi f_c t + (2i - 1) \frac{\pi}{4} \right] & 0 \leq t \leq T_s \\ 0 & \text{elsewhere} \end{cases} \quad (3.3)$$

where $i = 1, 2, 3, 4$, E is the transmitted symbol energy, T_s is the symbol duration and f_c is the carrier frequency. Each possible value of the phase corresponds to a unique pair of bits called a *dibit*. The mapping of blocks of two input bits to four modulation states is done by a code which permits adjacent phase states to differ by only one bit of the encoded two bit codeword. The code which permits this to occur is referred to as *Gray code*. The Gray encoded set of dibits are 10, 00, 01 and 11.

Fig 3 3(a), and (b) shows the modulator and demodulator for this scheme.

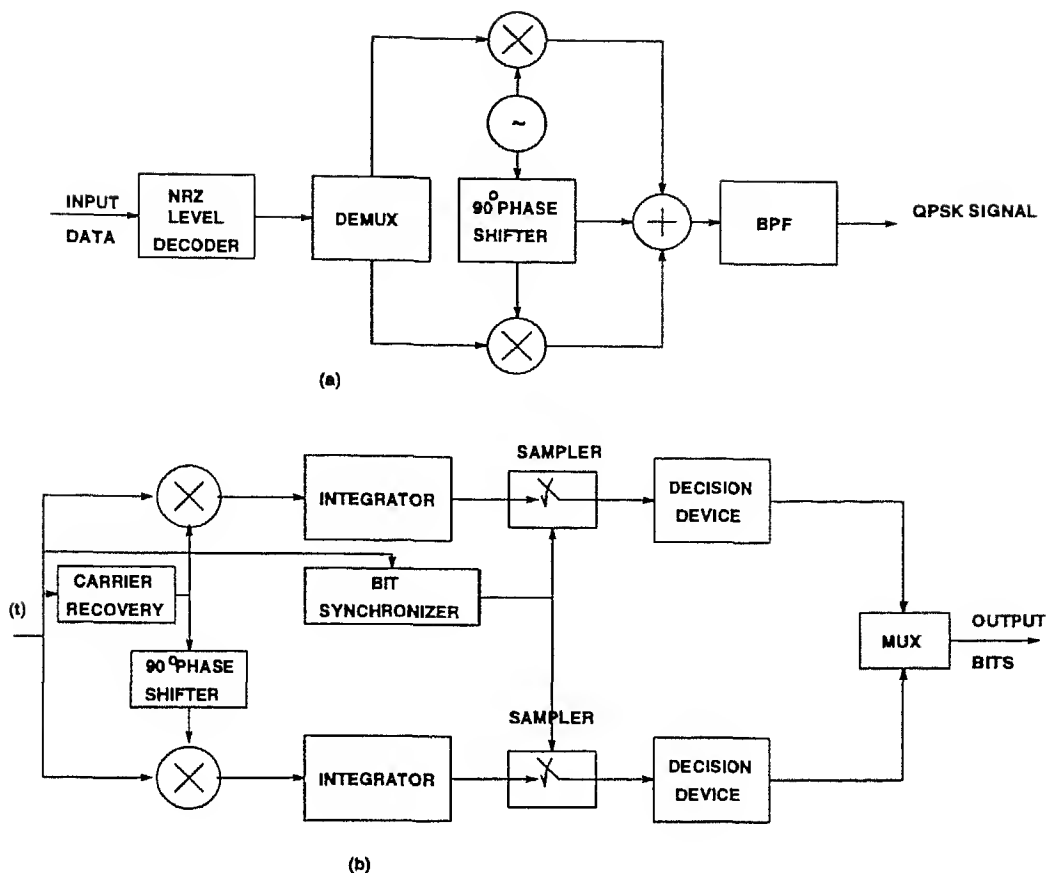


Figure 3.3 (a) Modulator, and (b) Demodulator for QPSK

3.1.3 Differential Phase - Shift Keying

The *differential phase shift keying* can be viewed as the noncoherent version of PSK. It eliminates the need for a coherent reference signal at the receiver by combining two basic operations at the transmitter:

- (1) *differential encoding* of the input binary waveform, and
- (2) *phase keying*, hence, the name, *differential phase shift keying* (DPSK).

The modulator and demodulator for DPSK are shown in Fig. 3.4(a), and (b).

The receiver is equipped with a storage capability, so that it can measure the *relative phase difference* between the waveforms received between two successive bit intervals.

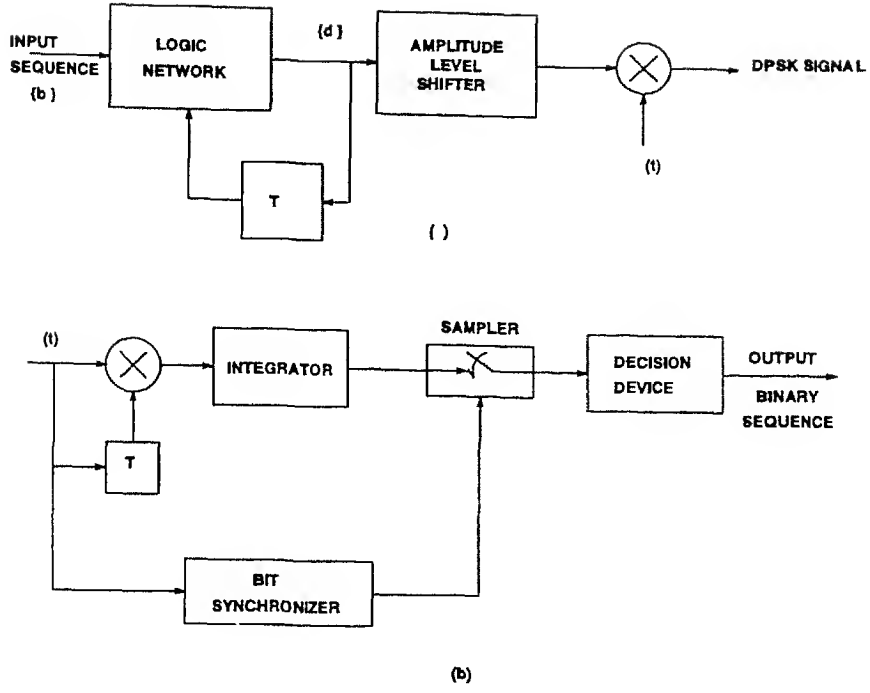


Figure 3.4 (a) Modulator, and (b) Demodulator for DPSK

The transmission of 1 leaves the carrier phase unchanged, so $s_1(t)$ is defined as

$$s_1(t) = \begin{cases} \sqrt{\frac{E_b}{2T_b}} \cos(\omega_c t + \theta_k) & kT_b \leq t \leq (k+1)T_b \\ \sqrt{\frac{E_b}{2T_b}} \cos(\omega_c t + \theta_k) & (k+1)T_b \leq t \leq (k+2)T_b \end{cases} \quad (3.4)$$

where $\theta_k = 0$ or π

The transmission of 0 advances the carrier phase by 180° so $s_2(t)$ is defined as

$$s_2(t) = \begin{cases} \sqrt{\frac{E_b}{2T_b}} \cos(\omega_c t + \theta_k) & kT_b \leq t \leq (k+1)T_b \\ \sqrt{\frac{E_b}{2T_b}} \cos(\omega_c t + \theta_k + \pi) & (k+1)T_b \leq t \leq (k+2)T_b \end{cases} \quad (3.5)$$

The generation of DPSK¹ signals is explained below

The differential encoding process at the transmitter input starts with an arbitrary first bit, serving as reference. Let d_k denote the differentially encoded sequence with this added reference bit. If the incoming binary symbol b_k is 1, d_k is left unchanged with respect to the previous bit, otherwise it is changed. The differentially encoded sequence d_k thus generated is used to phase shift a carrier with phase angles 0 and π corresponding to the bits in the sequence.

¹refer to Fig 3.4

3 1 4 Binary Frequency Shift - Keying

In a *binary frequency shift keying* BFSK the symbol 1 and 0 are distinguished from each other by transmitting one of the two sinusoidal waves that differ in frequency by a fixed amount. A typical pair of sinusoidal waves is defined by

$$s_i(t) = \begin{cases} \sqrt{\frac{2E_b}{T_b}} \cos[2\pi f_i t] & 0 \leq t \leq T_b \\ 0 & \text{elsewhere} \end{cases} \quad (3.6)$$

where $i = 1, 2$. $s_1(t)$ represents symbol 1 and $s_2(t)$ represents symbol 0.

$f_1 = f_c - \delta f$, $f_2 = f_c + \delta f$ where $f_c = \frac{f_1 + f_2}{2}$ is the nominal carrier frequency. As can be seen from equation 3.6, the switching between two frequencies f_1 and f_2 causes the phase of the two signals s_1 and s_2 to change abruptly; hence the system is not used in practice. When

$$s_i(t) = \begin{cases} \sqrt{\frac{2E_b}{T_b}} \cos[2\pi f_1 t + \theta(0)] & \text{for symbol 1} \\ \sqrt{\frac{2E_b}{T_b}} \cos[2\pi f_2 t + \theta(0)] & \text{for symbol 0} \end{cases} \quad (3.7)$$

it is a continuous phase signal in the sense that the phase continuity is always maintained including the inter-bit switching times. Here $\theta(0)$ denotes the values of phase at time $t = 0$ and it depends on the past history of modulation process. This form of modulation is an example of *continuous phase FSK* (CPFSK). As there are no abrupt phase changes in the signal, this results in a narrow spectrum. Hence it is generally used in practice. The generation and detection of CPFSK is shown in Fig. 3.5(a) and (b).

3 1 5 Minimum Shift Keying

CPFSK signal can also be represented as

$$s(t) = \sqrt{\frac{2E_b}{T_b}} \cos[2\pi f_c t + \theta(t)] \quad (3.8)$$

where $\theta(t) = \theta(0) \pm \frac{\pi h}{T_b} t$, $0 \leq t \leq T_b$

Here $h = T_b(f_1 - f_2)$ is referred to as the deviation ratio.

For $h = 2$

$f_1 - f_2 = \frac{1}{T_b}$. This is the *minimum frequency spacing* that allows the two FSK signals

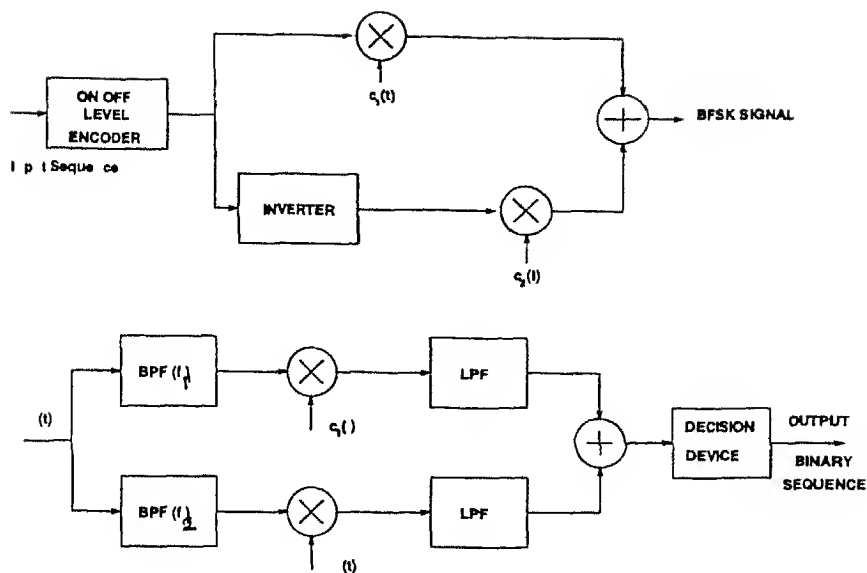


Figure 3.5 (a) Modulator, and (b) Demodulator for BFSK

representing symbols 1 and 0, not to interfere with one another in the process of detection. For this reason, such CPFSK signal is called *minimum shift keying*, MSK.

$s(t)$ in equation 3.8 can also be written as

$$s(t) = \sqrt{\frac{2E_b}{T_b}} \cos[\theta(t)] \cos 2\pi f_c t - \sqrt{\frac{2E_b}{T_b}} \sin[\theta(t)] \sin 2\pi f_c t$$

$$\theta(t) = \theta(0) \pm \frac{\pi}{2T_b} t, \quad 0 \leq t \leq T_b$$

where + sign corresponds to symbol 1 and - sign corresponds to symbol 0. Fig. 3.6(a) and (b) show the modulator and the demodulator for MSK.

3.1.6 M - PSK

In an M-PSK system, the phase of the carrier is permitted to be in any of the phase states $\phi_k = \frac{2\pi k}{M}$ for $k = 0, 1, 2, \dots, M-1$, each of these carrier states or signalling waveforms have equal energies E , and are transmitted by changing the *phase* of the carrier in M discrete steps. Here, M is 2^k and k , the number of bits in a symbol. The symbol rate R_s , of the coded signal used in the transmission, is reduced by the amount $\log_2 M$ from that of the binary case.

Hence, R_s is given by

$$R_s = \frac{R_b}{\log_2 M},$$

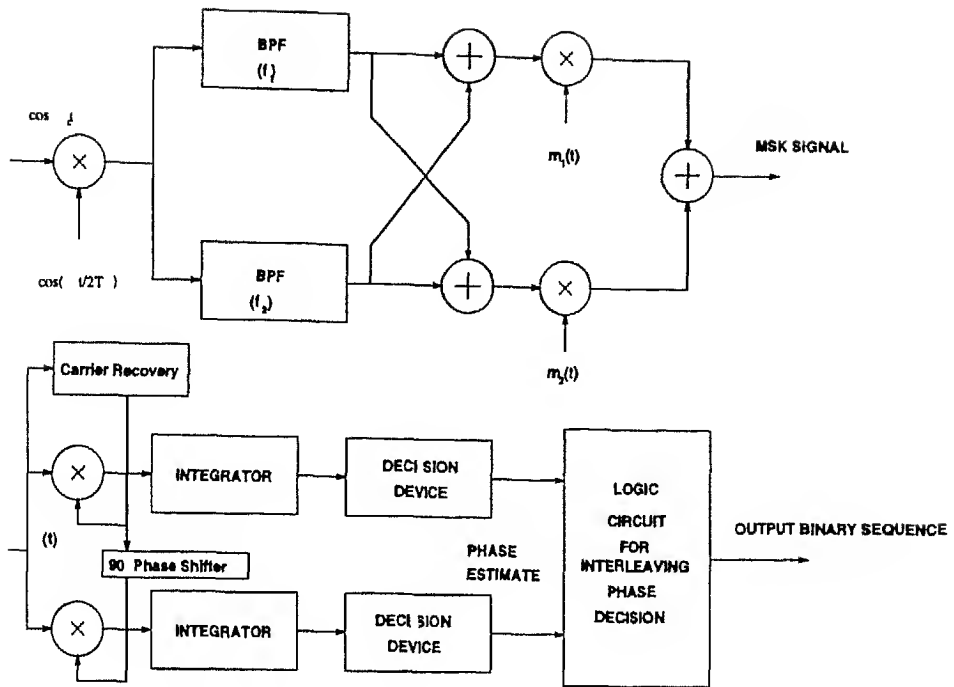


Figure 3 6 (a) Modulator and (b) Demodulator for MSK

If $M = 8$,

$$R_s = \frac{R_b}{3} \quad (\text{for } 8 \text{ PSK})$$

Thus, the 8 possible signals that can be transmitted in a symbol interval T_s ($T_s = \frac{1}{R_s}$) are given by

$$s_k(t) = \sqrt{\frac{2E}{T_s}} \cos(\omega_c t + \frac{2\pi k}{8}) \quad k = 0, 1, 2, \dots, 7 \quad (3.9)$$

The signal constellation for 8 PSK is shown in Fig 3 7 which shows that the signals in equation 3 9, correspond to the phase shifts of the carrier by $0^\circ, \pm 45^\circ, \pm 90^\circ, \pm 135^\circ, 180^\circ$

The mapping of blocks of three input signal bits to the various 8 modulation states is done by *Gray code*. Thus, in the demodulation process, because of the presence of noise the erroneous selection of adjacent phase states produces only a single bit error and not

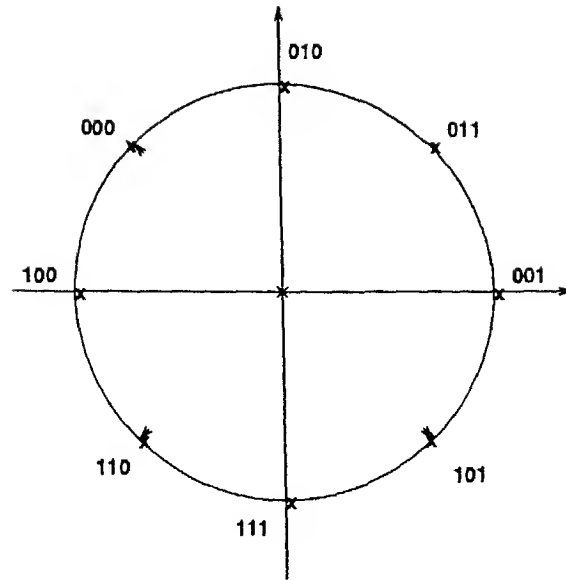


Figure 3.7 Signal Constellation for 8 PSK

a multiple bit error

In coherent detection, the best that the detector can do in the presence of *additive stationary white zero mean gaussian noise* is to make a guess of the transmitted message. Consequently, the measure of performance of the detector will be the number of times that it guesses wrongly in a long typical sequence of messages. A reasonable rule which has been adopted, assumes that the signal whose message point lies closest to the received signal is the message which is actually transmitted.

For coherent detection, the decision rule which selects the message point closest to the received point minimizes the probability of error. The probability of error, in terms of $\text{erfc}(u)$, the error complementary function² for the coherent case and for Gaussian noise having a *two sided power spectral density* of $N_0/2$, lies in a range as given below [3]

$$\frac{1}{2} \text{erfc} \left(\left[\frac{E_b}{N_0} \sin^2 \left(\frac{\pi}{M} \right) \right]^{1/2} \right) < P_e < \text{erfc} \left(\left[\frac{E_b}{N_0} \sin^2 \left(\frac{\pi}{M} \right) \right]^{1/2} \right)$$

² $\text{erf}(u) = \frac{2}{\sqrt{\pi}} \int_0^u \exp(-z^2) dz$ $\text{erfc}(u) = \frac{2}{\sqrt{\pi}} \int_u^\infty \exp(-z^2) dz$ Also $\text{erfc}(u) = 1 - \text{erf}(u) = 2Q(\sqrt{2}u)$
 where $Q(u) = \frac{1}{\sqrt{2\pi}} \int_u^\infty \exp(-\frac{z^2}{2}) dz$

Typically

$$P_e = \text{erfc} \left[\sqrt{\frac{2E_b}{N_o}} \sin \frac{\pi}{2M} \right] \quad (3.10)$$

3.2 Probability of error

Following expressions give the *probability of error* [4] and [5],

For BPSK

$$P_e = Q \left(\sqrt{\frac{2E_b}{N_o}} \right) \quad (3.11)$$

For QPSK

$$P_e = Q \left(\sqrt{\frac{2E_b}{N_o}} \right) \quad (3.12)$$

For DPSK

$$P_e = \frac{1}{2} \exp \left(-\frac{E_b}{N_o} \right) \quad (3.13)$$

For coherent BFSK

$$P_e = Q \left(\sqrt{\frac{E_b}{N_o}} \right) \quad (3.14)$$

For MSK

$$P_e = Q \left(\sqrt{\frac{2E_b}{N_o}} \right) \quad (3.15)$$

For 8 PSK

$$P_e = \text{erfc} \left[\sqrt{\frac{2E_b}{N_o}} \sin \frac{\pi}{16} \right] \quad (3.16)$$

3.3 Proposed Schemes for the System

3.3.1 For 6 GHz Case

At 6 GHz operating RF frequency, the allocated bandwidth, from Sec 1.4, is 30 MHz. Due to the bandwidth constraint³ BPSK, QPSK, BFSK, MSK modulation techniques cannot be used. So, for the 6 GHz case, the proposed digital modulation scheme is 8

PSK. In this, the phase of the carrier is permitted to be in any of the phase states $\phi_k = \frac{2\pi k}{8}$ for $k = 0, 1, 2, \dots, 7$, each of these carrier states or signalling waveforms

³For the bandwidth requirement of various schemes refer to Table 4.1 or 6.1

have equal energies E , and are transmitted by changing the *phase* of the carrier Fig 3 8 and 3 9 give the modulator and demodulator for 8 PSK The necessary *link budget calculations* are given in Chapter 4

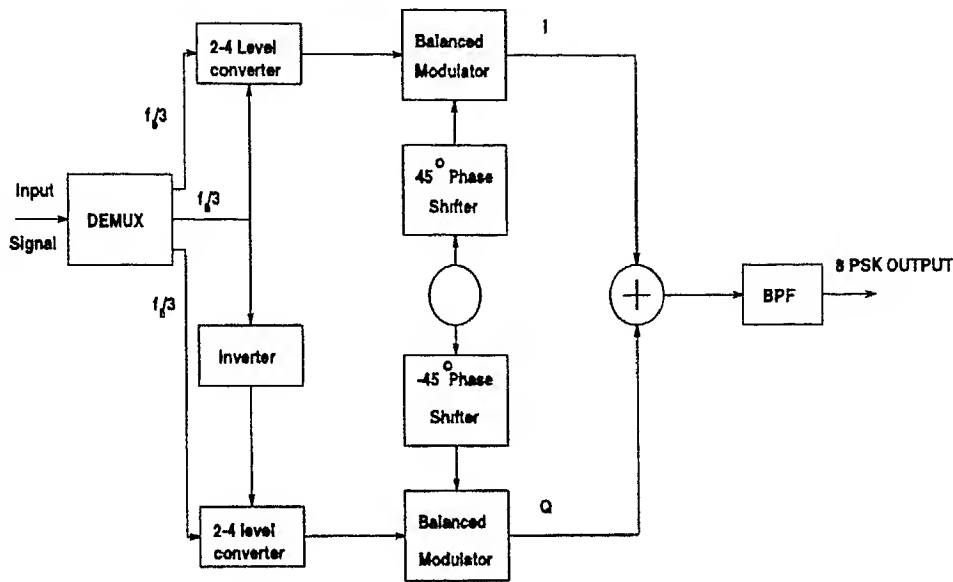


Figure 3 8 Modulator for 8 PSK

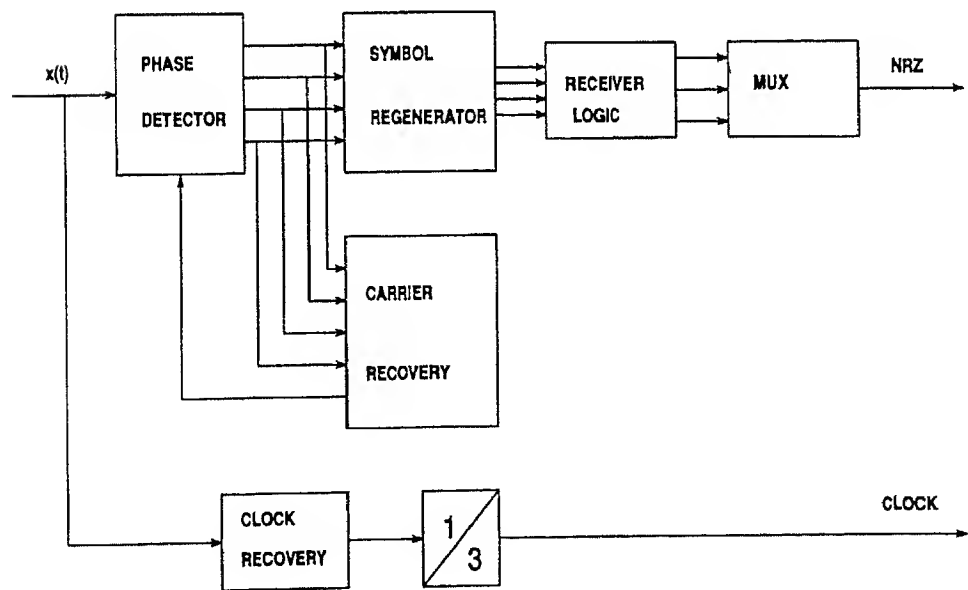


Figure 3 9 Demodulator for 8 - PSK

3 3 2 For 20 GHz Case

As mentioned in Sec 1 4, there is no bandwidth constraint for 20 GHz link so any of the schemes described in Sec 3 1 i.e BPSK QPSK, DPSK BFSK, or MSK can be used

3 4 Synchronization

The coherent reception of digitally modulated signal requires that the receiver be synchronous to the transmitter There is need for two basic modes of synchronization in a digital demodulator

- When coherent detection is used, knowledge of both the frequency and phase of the carrier is necessary The estimation of carrier phase and frequency is called *carrier recovery* or *carrier synchronization*
- To perform demodulation, the receiver has to know the starting and finishing instants of individual symbols, so that it may determine when to sample and when to quench the product integrators The estimation of these times is called *clock recovery* or *symbol synchronization*

3 4 1 Carrier Recovery

Fig 3 10 shows the block diagram of a carrier recovery circuit for M - ary PSK The circuit is called the *Mth power loop* For the binary case, the circuit is called a *squaring loop* The phase locked loop locks to double the carrier frequency, after dividing by two the coherent local carrier having the desired frequency is obtained However when the

Mth power loop is used for carrier recovery, we encounter a phase ambiguity problem Considering the squaring loop for BPSK, it is clear that because of the square device, sign change in input will leave the recovered signal unaltered, so there is a phase ambiguity of 180° Similarly, the M ary PSK exhibits M phase ambiguities in the interval $(0, 2\pi)$

Another method for carrier recovery involves the use of *Costas loop*, which consists of two paths, one referred to as *in phase* and other referred to as *quadrature* these two

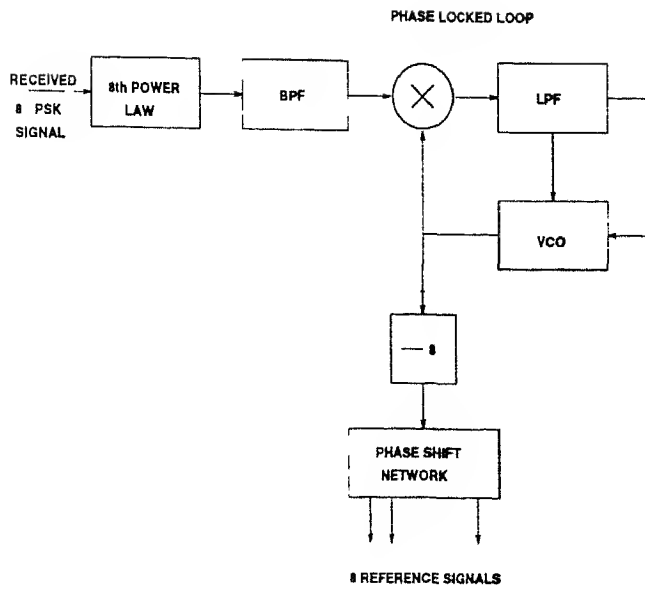


Figure 3 10 Mth Power Loop

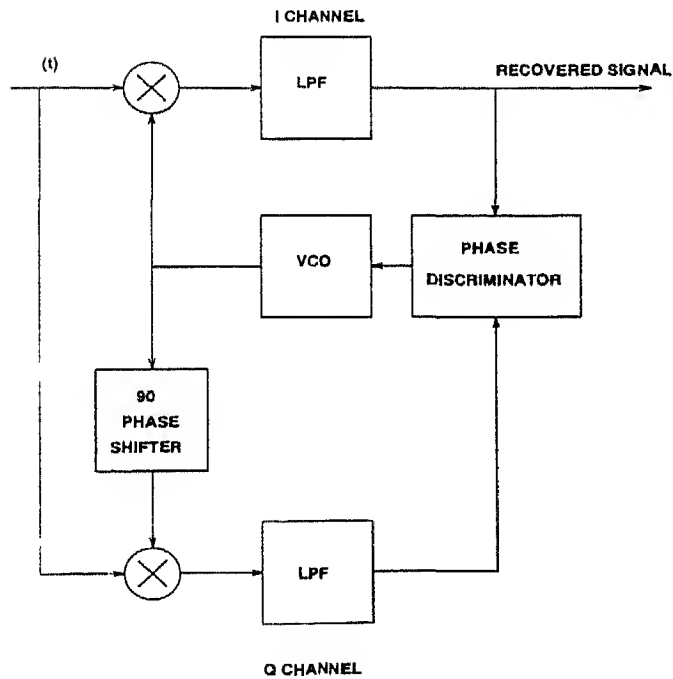


Figure 3 11 Costas loop

paths are coupled together via a common voltage controlled oscillator to form a negative feedback system⁴ When synchronization is attained the demodulated data waveform appears at the output of the in phase path

⁴refer to Fig 3 11

3 4 2 Symbol Recovery

A bit synchronizer circuit can be used to establish the times, at the receiver at which a bit interval ends and the succeeding bit interval begins. The synchronizer thus regenerates, at the receiver, a clock waveform which is synchronous with the original clock waveform, at the transmitter. A simple bit synchronizer, explained below can be used to extract clock from the signals having carrier to noise ratio of 6 dB or better [9]. The circuit is shown in Fig. 3.12. It is a phase locked loop in which the timing comparison is done by a *flip flop*.

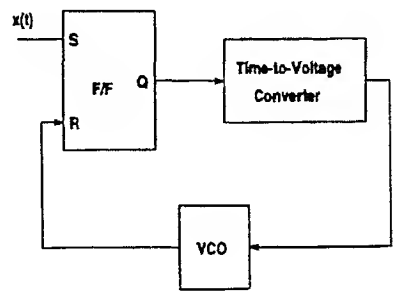


Figure 3.12 Bit Synchronizer

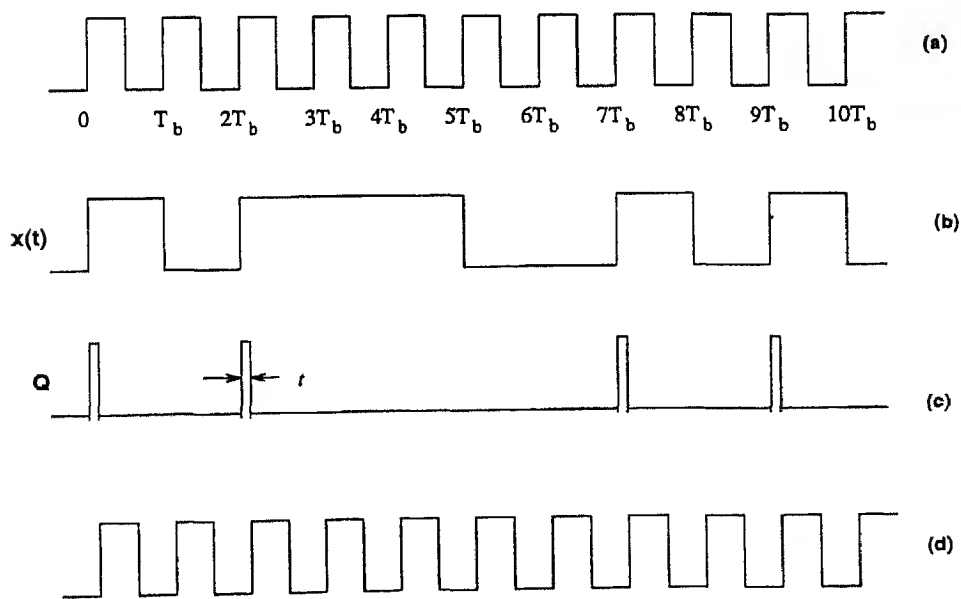


Figure 3.13 Waveforms of the Bit Synchronizer

The flip flop is designed to respond to the positive going *transitions*. Thus when the data $x(t)$ makes a positive transition from logic 0 to logic 1, the flip-flop will go to the

set state with $Q = 1$. A positive going transition on R(Reset) will transfer the flip-flop back to the reset state with $Q = 0$. Waveforms of the bit synchronizer are shown in Fig 3.13(a) to (d). The original timing clock of the transmitter is shown in (a) and the recovered clock is shown in (c). The *time to voltage converter* shown in the figure is a circuit which on receiving an input pulse from the flip-flop generates a dc voltage which is proportional to the pulse duration t and then holds its output v_o constant until updated on the arrival of next received pulse. Now if $t = T_b/2$ the converter output v_o is exactly the voltage which causes the VCO to oscillate at the correct clock frequency, i.e. at the data bit rate $f_b = 1/T_b$.

Chapter 4

Link Budget Calculations

Link budget calculations involve power budget, noise budget and calculation of power to be transmitted The minimum receivable power is the starting point of system design

A microwave link can span a distance of a few kilometers to several thousand kilometers Each hop is surveyed for a line of sight antenna path having necessary clearance The size of the antennas transmitter output power, minimum acceptable receive power and hop length are all interrelated

4.1 CASE I 6 GHz

4.1.1 Carrier to Noise Power Ratio $\frac{C}{N}$

The probability of error¹ is given by

$$P_e = \text{erfc} \left(\sqrt{\frac{2E_b}{N}} \sin \frac{\pi}{2M} \right) \quad \text{where } M = 8 \text{ So}$$

$$\frac{E_b}{N_0} = 22.693 \text{ dBHz for a } P_e \text{ of } 10^{-7} \quad [4, 5]$$

Adding degradation, (refer to table 2.2), it becomes

$$\frac{E_b}{N_0} = 26.893 \text{ dBHz}$$

The carrier to noise density ratio $\frac{C}{N}$ is given by,

$$\frac{C}{N} = \frac{E_b}{N_0} R_b \quad (\text{ratio})$$

¹Refer to section 3.2 of Chapter 3

where $R_b = 34 \text{ Mbps}$

Hence

$$\frac{C}{N} = 102\,208 \text{ dBHz}$$

Noise bandwidth for 8 PSK from table 4.1 is

$$BW = 28.33 \text{ MHz}$$

Carrier to noise ratio is,

$$\frac{C}{N} = \frac{C}{N} - BW_{(\text{dB})} \text{ dB}$$

Hence

$$\frac{C}{N} = 27\,685 \text{ dB} \quad (4.1)$$

4.1.2 Noise Power N

The input noise temperature of the receiving system is given by

$$T_{\text{in}} = G_1 T_A + T_p(1 - G_1)$$

where,

$$G_1 = 4 \text{ dB the waveguide loss}$$

$$T_A = 140 \text{ K, antenna noise temperature,}$$

$$T_p = 290 \text{ K, ambient temperature}$$

The system noise temperature is given by,

$$T_{\text{sys}} = T_{\text{in}} + (NF - 1)T_p G_1$$

where NF is the noise figure(ratio) of the receiver

In the present case,

$$T_{sys} = 3591.2 \text{ K}$$

Noise Power is defined as

$$N = kTB \quad \text{watts, where}$$

$$k = -228.6 \text{ dB}, \text{ the Boltzmann's constant}$$

$$T = T_{sys} \text{ and,}$$

$$B = BW$$

So the noise power is,

$$N = -228.6 + 10 \log_{10} T_{sys} + 10 \log_{10} BW \quad \text{dBW}$$

$$N = -118.48 \quad \text{dBW} \quad (4.2)$$

4.1.3 Power to be Transmitted P_t

From equations 4.1 and 4.2

$$C = P_r = -90.7925 \quad \text{dBW} \quad (4.3)$$

where, P_r is the received power at the receiver input

$$P_t = P_r + \sum L - \sum G \quad \text{dBW} \quad (4.4)$$

where P_t is the power to be transmitted

From equations 4 3 2 2 2 10 the required transmitted power is

$P_t = 5.7275 \text{ dBm} = 3.74 \text{ mW}$ (4 5)

for dry mountain terrain

4 2 Case II 20 GHz

By following the steps in Sec 4 1 and using equations 3 11 to 3 15, 2 11 2 3 the link budget calculations are tabulated in the Table 4 1

Scheme	$\frac{E_b}{N}$ dBHz	$\frac{C}{N}$ dBHz	BW MHz	$\frac{C}{N}$ dB	N dBW	P_r dBW	P_t^\dagger dBm	P_t^\ddagger dB
BPSK	15 49	90 80	85	11 5	113 68	102 18	2 92	2 92
QPSK	15 49	90 80	35 78	15 26	117 42	102 16	2 94	2 94
DPSK	16 38	91 69	35 78	16 15	117 42	101 27	3 83	3 83
MSK	15 49	90 80	57 63	13 19	-115 37	102 18	2 92	2 92
BFSK	19 65	94 96	85	15 67	113 68	98 01	7 09	7 09

* Practical values from [2] † for moderate rainfall ‡ for heavy rainfall

Table 4 1 Link Budget Calculations for different modulation schemes

Chapter 5

System Description

The system has to transmit the modulated T V signal from the studio end to the transmitter end, from where after necessary demodulation and decoding it can be broadcast on a terrestrial link. A brief description of the system using a block schematic is given in this chapter alongwith some important specifications of the transmitter and the receiver.

5.1 Brief Description of the Subsystems

The overall block schematic of the system is given in Fig. 5.1. The input of the system is the encoded signal from studio. *Here it is assumed that the sampling, analog to digital conversion and the necessary coding of the T V signal in the standard format¹ is already done in the studio.* The transmitter part of the system could be installed in the *master switching room* of the studio, just like the analog transmitter. Line code conversion, error coding and signal predistortion are some of the important processes that should take place before the signal is available at the input of the transmitter. The line code converter takes the standard PCM² line codes and converts the code to NRZ format. The resulting code is scrambled by means of a PRBS³. This action tends to remove internal correlation among symbols such as long string of 1s or 0s. Error coding

¹As per the recommendations of CCIR Rec. 601

²8 bit pulse code modulated

³Pseudo Random Binary Sequence

is done for the forward error correction in the transmission as the bits corresponding to synchronization should be decoded correctly. The predistorter is designed to compensate for the distortion imparted on the signal by the power amplifier.

This baseband signal is to be modulated by digital modulator to give 70 MHz IF using the schemes proposed in Chapter 3. The output of the modulator is then amplified, filtered and passed to the upconverter. The upconverter translates this signal to the operating frequency of the system. The output of the upconverter is then fed to the RF amplifier⁴ which amplifies the signal to the desired output level. The output of the amplifier is passed through a bandpass filter to reduce the spurious and harmonic out of band signals.

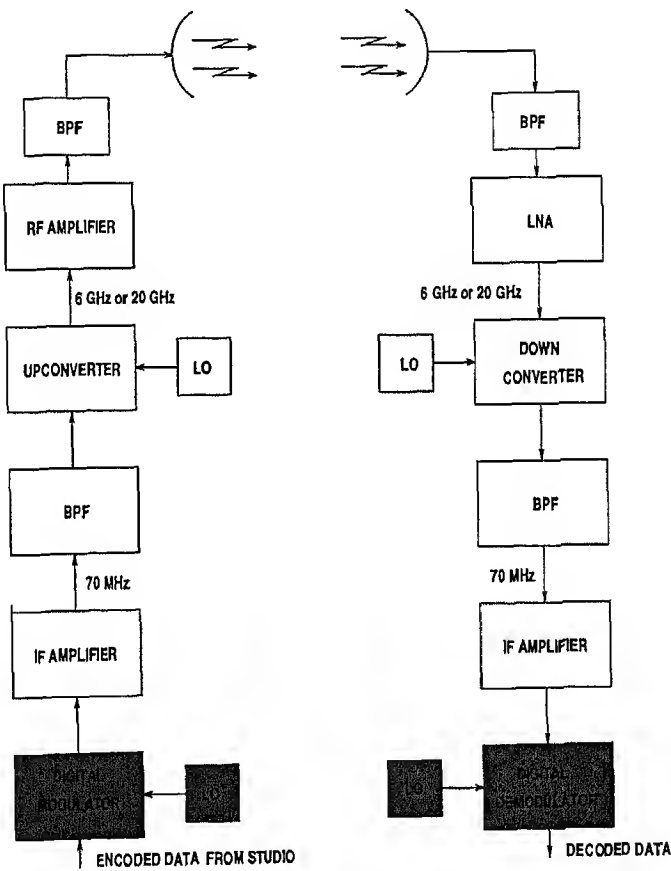


Figure 5.1 Overall System Diagram

⁴ A microwave tube amplifier like TWT or solid state microwave amplifier

This signal can now be fed to the antenna subsystem via the waveguide run radiation to the distant end

The received signal, starting from the antenna downward in Fig 5 1 is fed from antenna subsystem, via the waveguide run, through a bandpass filter through the L to the downconverter. Commonly available demodulators have noise figure ranging from 7 dB to 12 dB. The down converter translates the incoming signal to the IF which is 70 MHz. The output of the downconverter is then fed through a bandpass filter to the IF amplifier prior to feeding into the digital demodulator. The demodulation techniques chosen here are coherent type, so a carrier recovery system is also required. This system is explained in chapter 3. Except for the blocks shown in gray colour in Fig 5 1, remaining subsystems can be used from the existing analog system.

5 2 Specifications of the System

5 2 1 Transmitter

- 1 Radio Frequency Band 6 GHz
- 2 Channel Provision One Video & One Audio
- 3 Stability of RF local oscillator $\pm 5 \times 10^{-6}$
- 4 Transmitter Amplifier Output Refer to table 4 1 for P_t
- 5 Transmitter IF Frequency 70 MHz \pm 10 KHz
- 6 IF output Level +5 2 dBm \pm 0 5dB
- 7 T V System Applicable CCIR System B
- 8 Carrier Frequency for modulator 70 MHz

⁵Low Noise Amplifier

9 Modulator Specifications

- Input level ECL compatible
- Bit rate 34 Mbps
- Pulse shape Raised cosine
- Bandwidth Refer to table 4.1 for BW

5.2.2 Receiver

- 1 Radio Frequency Band 6 GHz
- 2 Channel Provision One Video & One Audio
- 3 Stability of RF local oscillator $\pm 5 \times 10^{-6}$
- 4 Receiver input level Refer to Table 4.1 for P_r
- 5 Noise figure (at the receiver input) 11 dB
- 6 Intermediate Frequency 70 MHz
- 7 IF output Level 5.2 dBm ± 0.5 dB
- 8 A G C Range + 5 dB to 40 dB
- 9 T V System Applicable CCIR System B

10 Demodulator Specifications

- Output level ECL compatible
- Bit rate 34 Mbps
- Bandwidth Refer to table 4.1 for BW

Chapter 6

Results and Conclusions

Table 6.1 gives a comparative study of various schemes. It can be seen that for efficient bandwidth utilization, many techniques are suitable, hence for 6 GHz case the only choice is 8 PSK. But when there is no bandwidth constraint coherent BPSK, coherent QPSK, or DPSK could be used. BFSK requires comparatively large amount of power, so it is not suitable. However, MSK is suitable as power requirement is small in this case. Considering the bandwidth efficiency and the power requirement together QPSK can be chosen for 20 GHz case. Moreover, as all schemes discussed in this thesis are coherent type, the circuit complexity is not an important factor while choosing a particular scheme for 20 GHz case. The table gives P_t for dry mountain terrain case and moderate rainfall.

Frequency	Scheme	$\frac{C}{N}$ dB	BW MHz	P_t dBm
6 GHz	8 PSK	27.685	28.33	5.73
20 GHz	BPSK	11.5	85	2.92
20 GHz	QPSK	15.26	35.78	2.94
20 GHz	DPSK	16.15	35.78	3.83
20 GHz	BFSK	15.67	85	7.09
20 GHz	MSK	13.19	57.63	2.92

Table 6.1 Comparative Study of the Schemes

6 1 Suggestions for Further Work

Further work can be taken up in the following areas

- The subsystem design for the systems proposed in this thesis
- Digital system and subsystem design for the analog DOT links that are presently used by Doordarshan
- Design of the system when adjacent channel interference is to be considered
- The system design for a digital satellite link
- Study and system design for terrestrial broadcast of digital video in VHF and UHF bands with *high definition T V* (HDTV)

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